SYSTEM AND METHOD FOR DATA COMMUNICATION

OVER MULTI-INPUT, MULTI-OUTPUT CHANNELS

5   *FIELD OF THE INVENTION*

The present invention relates generally to data transmission systems and, more particularly,

to transmission and detection techniques for use in an environment characterized by a single

or several multi-input, multi-output channels.

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BACKGROUND OF THE INVENTION

Reliable wireless transmission is difficult to achieve, since the time varying nature of

multipath fading channels makes it a challenge when compared to communication over

15   optical fibre, coaxial cable and copper twisted pair. Although an improvement in bit error

rate (BER) can be achieved by increasing the power or bandwidth of transmitted signals,

this would be contrary to the efficiency requirements of next generation wireless systems.

Diversity schemes, based on multiple transmissions/receptions of data packets, have

therefore been proposed as techniques to mitigate fading without increasing the transmitted

20   power.

Prior diversity schemes have included frequency and time diversity. However, these

approaches have the disadvantage that additional spectrum or time slots are used, thus

reducing the overall system efficiency. On the other hand, space diversity, in which

25   multiple transmit and receive antennas allow separation of received signals using spatial

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processing techniques, remains a possible alternative allowing improved overall economy

and efficiency of a digital cellular system. Space diversity techniques based on multiple

transmit and receive antennas are commonly referred to as MIMO (multi-input, multi-

output) systems.

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However, the various MIMO systems in existence typically trade off implementational

simplicity against total achievable data rate. In other words, prior MIMO systems either

achieve a high data rate at the expense of either high power consumption or high hardware

complexity, or maintain low transmit power but provide a correspondingly low data rate. In

10   either case, co-antenna interference (CAI) is the dominant limiting factor in achieving high

data rates. This trade off is a drawback when designing for wireless transmission, since low

power consumption and high data rate are both desirable. Hence, there is a need in the

industry to provide a high-rate transmission and detection scheme, which can be

implemented with relatively low complexity both at the transmitter and the receiver.

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SUMMARY OF THE INVENTION

The present invention may be summarized according to a broad aspect as a system for

estimating data transmitted by a plurality of transmit elements across a communications

20   channel. The system comprises a plurality of receive interfaces, each receive interface

operative to receive a signal via the communications channel and output a respective

sequence of received data elements. The system also comprises a space matched filter

connected to the plurality of receive interfaces and operative to (i) assemble the received

data elements into sets of received data elements, each the set of received data elements

25   including at least one received data element from each sequence of received data elements;

and (ii) jointly process each set of received data elements with each of a plurality of channel

data elements to produce a corresponding plurality of filtered data elements, each filtered

data element being associated with one of the transmit elements, each channel data element

being representative of a portion of the communications channel between an associated one

5   of the transmit elements and the plurality of receive interfaces. Finally, the system

comprises a detector connected to the space matched filter and operative to process each

filtered data element to produce a corresponding decision data set therefor.

According to a second broad aspect, the present invention seeks to provide a method for

10   estimating data transmitted by a plurality of transmit elements across a communications

channel. The method comprises receiving a plurality of signals at a plurality of receive

interfaces and outputting a respective plurality of sequences of received data elements. The

method also comprises assembling the received data elements into sets of received data

elements, each set of received data elements including at least one received data element

15   from each sequence of received data elements. The method further comprises jointly

processing each set of received data elements with each of a plurality of channel data

elements to produce a corresponding plurality of filtered data elements, each filtered data

element being associated with one of the transmit elements, each channel data element

being representative of a portion of the communications channel between an associated one

20   of the transmit elements and the receive interfaces. The method also comprises processing

each filtered data element to produce a corresponding decision data set therefor.

In accordance with a third embodiment of the present invention, there is provided a

computer-readable storage medium containing a program element for execution by a

25   computing device to implement a space matched filter. The space matched filter comprises

a control entity; and an I/O entity for receiving an array of data elements including at least

one data element associated with each of a plurality of receive interfaces in communication

with a plurality of transmit elements across a communications channel. The control entity is

operative to jointly process each array of received data elements with each of a plurality of

5 channel data elements to produce a corresponding plurality of filtered data elements, each

filtered data element being associated with one of the transmit elements, each channel data

element being representative of a portion of the communications channel between an

associated one of the transmit elements and the plurality of receive interfaces.

10 The present invention can also be summarized according to a fourth broad aspect as a

system for data communication over a multi-input, multi-output (MIMO) channel. The

system comprises a transmitter unit including a de-multiplexer for separating an information

stream into a plurality of information sub-streams and a plurality of transmit interfaces for

simultaneously transmitting respective ones of the information sub-streams over the MIMO

15 channel. The system also comprises a receiver unit, including a plurality of receive

interfaces for receiving respective received signals via the MIMO channel and transforming

the received signals into streams of received data elements; a space matched filter connected

to the receive interfaces, the space matched filter adapted to jointly process a vector formed

from received data elements from each stream of received data elements together with each

20 of a plurality of channel data elements, each channel data element being representative of a

portion of the MIMO channel between an associated one of the transmit interfaces and the

plurality of receive interfaces; a detector connected to the space matched filter and operative

to process each filtered data element to produce a corresponding decision data set therefor;

and a multiplexer for combining multiple decision data sets for filtered data elements

associated with different ones of the transmit interfaces into a single sequence of decision

data sets.

The present invention can also be summarized according to a fifth broad aspect as a system

5   for estimating data transmitted from each of a plurality of users across a shared

communications channel, the data transmitted from the kth user being transmitted via n_k

respective transmit elements. The system comprises a plurality of receive interfaces, each

receive interface operative to receive a signal via the communications channel and output a

respective sequence of received data elements. The system also comprises a space matched

10   filter connected to the plurality of receive interfaces and operative to (i) assemble the

received data elements into sets of received data elements, each set of received data

elements including at least one received data element from each sequence of received data

elements; and (ii) for each of the plurality of users, jointly process each set of received data

elements with each of a plurality of channel data elements to produce a corresponding

15   plurality of filtered data elements for the user, each filtered data element being associated

with one of the transmit elements for the user, each channel data element being

representative of a portion of the communications channel between an associated one of the

transmit elements for the user and the plurality of receive interfaces. The system further

comprises a detector connected to the space matched filter and operative to process, for each

20   user, each filtered data element for the user to produce a corresponding decision data set

therefor.

These and other aspects and features of the present invention will now become apparent to

those of ordinary skill in the art upon review of the following description of specific

25   embodiments of the invention in conjunction with the accompanying drawings.

BRIEF DESCRIPTION OF THE DRAWINGS

In the accompanying drawings:

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Fig. 1 shows, in block diagram form, a transmitter in accordance with a single-user

embodiment of the present invention;

Figs. 2, 3 and 4 show, in block diagram form, variants of a receiver in accordance with a

10   single-user embodiment of the present invention;

Fig. 5 shows, in block diagram form, a transmitter in accordance with a multi-user

embodiment of the present invention; and

15   Fig. 6 shows, in block diagram form, options for a receiver in accordance with a multi-user

embodiment of the present invention.

DETAILED DESCRIPTION OF THE EMBODIMENTS

20   Fig. 1 shows a transmitter 20 in accordance with an embodiment of the present invention.

The transmitter 20 receives information symbols 10 constituting a message from a digital

data source 22. The information symbols 10 are first passed through a forward error

correction (FEC) encoder 24, which is optional. The FEC encoder 24, if used, includes

suitable circuitry, software and/or control logic for adding redundancy and/or memory to the

25   information symbols 10 by encoding them into a sequence of coded symbols 12. Suitable

embodiments of the FEC encoder 24 include but are not limited to processing blocks

implementing encoding operations such as block encoding (e.g., Reed-Solomon encoding),

convolutional encoding, parallel or serial concatenated codes (turbo codes based on the

concatenation of finite state machine (FSM) encoders) and so on.

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The coded symbols 12 are fed to a mapper 26, which includes suitable circuitry, software

and/or control logic for translating the coded symbols 12 into a stream of modulation

symbols 30. For example, in a specific non-limiting example embodiment, the coded

symbols 12 are bits and the modulation symbols are groups of $\log_2 M$ bits, where M is the

10   number of possible modulation symbols. Each modulation symbol includes $\frac{1}{2}\log_2 M$ coded

symbols mapped to a component in a first dimension (having $\sqrt{M}$ possible values) and

$\frac{1}{2}\log_2 M$ coded symbols mapped to a component in a second dimension (also having $\sqrt{M}$

possible values). The resulting modulation symbol thus can be viewed as representing one

of M points in a two-dimensional constellation.

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The stream of modulation symbols 30 is fed to a de-multiplexer 32, which includes suitable

circuitry, software and/or control logic for separating the modulation symbols 30 into a

plurality of modulation symbol sub-streams 34. The de-multiplexer 32 is connected to a

channel interleaver 36, which includes suitable circuitry, software and/or control logic for

20   changing the order in which modulation symbols appear within each of the modulation

symbol sub-streams 34. The output of the channel interleaver 36 is therefore also a plurality

of modulation symbol sub-streams 38. The effect of the channel interleaver 36 is to

increase the degree to which channel fading is independent between successive symbols in

each of the modulation symbol sub-streams 38. One possible implementation for the

25   channel interleaver 36 is a buffer, which includes a two-dimensional array (i.e., a matrix)

for each modulation symbol sub-stream 38 that is written to in a row-by-row manner and read from in a column-by-column manner.

The modulation symbol sub-streams 38 produced by the channel interleaver 36 are fed to a

5   respective plurality of transmit antennas $40_1, 40_2, \ldots, 40_{n_T}$ via a discrete-time (DT) to continuous-time (CT) unit 42. The DT-to-CT unit 42 comprises suitable circuitry, software and/or control logic for applying a suitable modulation scheme to transform each modulation symbol in each of the modulation symbol sub-streams 38 into a continuous-time waveform. Specifically, the output of the DT-to-CT unit 42 is a plurality of continuous-

10   time waveforms 44, which are then sent across a wireless channel 46 by the plurality of transmit antennas $40_1, 40_2, \ldots, 40_{n_T}$. It will be noted that the number of transmit antennas $40_1, 40_2, \ldots, 40_{n_T}$ (and the number of digital symbol sub-streams 38) is n_T, where n_T is an integer greater than or equal to two. This permits efficient reuse of the transmission spectrum. The transmit antennas $40_1, 40_2, \ldots, 40_{n_T}$ can be of any suitable type, including

15   but not limited to omni-directional, sectorized, polarization, micro-array, adaptive, smart and so on.

Examples of suitable modulation schemes that can be used by the DT-to-CT unit 42 include, but are not limited to, binary phase-shift keying (BPSK), quadrature phase-shift keying

20   (QPSK), pulse amplitude modulation (PAM), quadrature amplitude modulation (QAM) and spread spectrum modulation (used in code division multiple-access (CDMA) based systems). For example, in M-ary QAM, each digital symbol consists of two components, one of which governs the amplitude of a given carrier signal, while the other governs the amplitude of a phase-shifted version of that carrier signal.

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For notational convenience, it is useful to refer to the collection of individual modulation

symbols, one from each of the modulation symbol sub-streams 38 (transmitted at a given

symbol instant t), as a "transmit symbol vector", hereinafter denoted $\mathbf{b}_t = \left[b_{t,1}, b_{t,2}, ..., b_{t,n_T} \right]^T$,

where ^{T} is the transpose operation. In one embodiment, the modulation symbols in the

5   transmit vector $\mathbf{b}_t$ will have the same transmission period (i.e., the corresponding

continuous-time waveform lasts the same amount of time), will be transmitted

simultaneously by the transmit antennas 40_1, 40_2, ..., 40_{n_T} and will represent the same

number of coded symbols. Of course, those skilled in the art will find it within their

capabilities to modify these parameters in order to suit particular design requirements.

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While traveling through the wireless channel 46, the signals transmitted by the transmit

antennas 40_1, 40_2, ..., 40_{n_T} suffer from co-antenna interference (CAI). Hence, the signals

at each of n_R receive antennas will be a superposition of the various transmitted signals. For

mathematical convenience, let it be assumed that signals traveling across the wireless

15   channel 46 are affected by quasi-static Rayleigh fading and additive white Gaussian noise,

that there is no line of sight (LOS) between any of the transmit antennas and any of the

receive antennas. In the case of Rayleigh fading, the amplitudes of the path gains are

modeled as samples of a complex Gaussian random variable with a mean of zero and a

variance of 0.5 in each dimension. It should be understood, however, that these

20   assumptions are made for mathematical convenience only, and in no way constitute an

operational requirement of the present invention. In fact, the present invention will provide

improvements regardless of whether the fading is Ricean, Rayleigh, etc., and, moreover,

regardless of whether the fading is correlated or uncorrelated.

Based on the above-described channel model, the received vector $\mathbf{r}_t = \left[r_{t,1}, r_{t,2}, ..., r_{t,n_R} \right]^T$ at

the t-th symbol instant for a transmit symbol vector $\mathbf{b}_t = \left[b_{t,1}, b_{t,2}, ..., b_{t,n_T} \right]^T$ will be given by:

$$\mathbf{r}_t = \mathbf{Hb}_t + \mathbf{v}_t,$$

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where $\mathbf{H}$ denotes a channel transfer matrix used to describe the effect of fading and multi-

antenna interference on the multi-input, multi-output (MIMO) channel 46 and $\mathbf{v}_t$ is additive

noise (often assumed to be additive white Gaussian noise (AWGN)). Specifically:

$$\mathbf{H} = \begin{bmatrix} h_{1,1} & h_{1,2} & \cdots & h_{1,n_T} \\ h_{2,1} & h_{2,2} & \cdots & h_{2,n_T} \\ \cdots & \cdots & \cdots & \cdots \\ h_{n_R,1} & h_{n_R,2} & \cdots & h_{n_R,n_T} \end{bmatrix},$$

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where coefficient $h_{i,j}$ is the path gain from the j-th transmit antenna to the i-th receive

15   antenna, $1 \leq j \leq n_T$, $1 \leq i \leq n_R$.

With reference to Fig. 2, the general structure of a receiver 60 in accordance with an

embodiment of the present invention is now described. The receiver 60 includes a plurality

of receive antennas $62_1, 62_2, ..., 62_{n_R}$. It is envisaged that $n_R \geq n_T$. This is usually feasible

20   in an uplink communication scenario because an abundance of antennas and processing

power will usually be permitted at the receiver. However, it should be understood that the

present invention is not restricted to the uplink communication scenario, nor is there a

requirement that for a specific relationship between n_T and n_R. In general, all that is

required is that n_T and n_R, the numbers of transmit and receive antennas, be greater than one.

25   This is generally achievable in both upstream and downstream scenarios.

The receive antennas 62_1, 62_2,..., 62_{n_R} can be of any suitable type, including but not

limited to omni-directional, sectorized, polarization, micro-array, adaptive, smart and so on.

The output of each of the n_R receive antennas 62_1, 62_2,..., 62_{n_R} is a respective continuous-

time waveform 64_1, 64_2,..., 64_{n_R}, which is provided to a respective portion of a CT-to-DT

5   unit 66. The CT-to-DT unit 66 comprises suitable circuitry, software and/or control logic

for discretizing each continuous-time waveform to produce a sequence of numerical values

therefrom. In particular, the CT-to-DT unit 66 is operative to demodulate the continuous-

time waveforms 64_1, 64_2,..., 64_{n_R} using a demodulation scheme that is the opposite of the

modulation scheme used by the DT-to-CT unit 42.

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For example, in the case of QAM, the CT-to-DT unit 66 is operable to determine the

amplitude of a first component that appears to be modulating a predetermined carrier and

the amplitude of a second component that appears to be modulating the 90-degree phase-

shifted version of that carrier. For notational convenience, both of these components are

15   represented in a single complex number. Hence, the elements of r_t can be real or complex

values.

The vector r_t emerging from the DT-to-CT unit 66 is provided to a space matched filter 68.

The basic idea of the space matched filter 68 is to correlate (or match) the vector r_t with a

20   "vector of chip amplitudes" corresponding to each of the transmit antennas 40_1, 40_2,...,

40_{n_T}. The "vector of chip amplitudes" corresponding to a given transmit antenna is

indicative of a channel characteristic associated with that transmit antenna. In one

embodiment, the "vector of chip amplitudes" corresponding to a given transmit antenna is

indicative of the degree to which the signal from that transmit antenna has been spread out

amongst the receive antennas $62_1, 62_2, \ldots, 62_{n_R}$.

In the context of equalization of a multi-input multi-output channel such as the wireless

5   channel 46, the "vector of chip amplitudes" corresponding to the jth transmit antenna is

ideally the jth normalized column of the channel transfer matrix $\mathbf{H}$, which is denoted s_j. In

other words, $\mathbf{H}$ can be rewritten as $\mathbf{H} = \mathbf{SA}$, where:

S is the matrix of "vectors of chip amplitudes", i.e.:

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\begin{aligned}
\mathbf{S} \quad &= \quad [\mathbf{s}_1 \ \ldots \ \mathbf{s}_{n_T}] \\[2mm]
&= \quad [\mathbf{h}_1/\|\mathbf{h}_1\| \ \ldots \ \mathbf{h}_{n_T}/\|\mathbf{h}_{n_T}\|] \\[2mm]
&= \quad \begin{bmatrix} s_{1,1} & \cdots & s_{n_T,1} \\ \cdots & \cdots & \cdots \\ s_{1,n_R} & \cdots & s_{n_T,n_R} \end{bmatrix}
\end{aligned}
$$

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and

A is an amplitude matrix, i.e.:

$$
\begin{aligned}
\mathbf{A} \quad &= \quad [\|\mathbf{h}_1\| \ \ldots, \|\mathbf{h}_{n_T}\|] \\[2mm]
&= \quad \begin{bmatrix} A_1 & \cdots & 0 \\ \cdots & \cdots & \cdots \\ 0 & \cdots & A_{nT} \end{bmatrix}
\end{aligned}
$$

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25   Therefore, space matched filtering performed on the received vector r_t yields a filtered

vector, denoted $\mathbf{y}_t = \left[y_{t,1}, y_{t,2}, \ldots, y_{t,n_T} \right]^T$ and represented as follows:

(Equation 1A)

$$\mathbf{y}_t = \mathbf{S}^H \mathbf{r}_t.$$

Now, it has already been shown that $\mathbf{r}_t = \mathbf{Hb}_t + \mathbf{v}_t$. Therefore, since $\mathbf{H} = \mathbf{SA}$, one has:

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$$\mathbf{r}_t = \mathbf{SAb}_t + \mathbf{v}_t.$$

Therefore, the filtered vector $\mathbf{y}_t$ from (Equation 1A) can be expressed in terms of a

correlation matrix $\mathbf{R}$:

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(Equation 1B)

$$\mathbf{y}_t = \mathbf{S}^H \mathbf{SAb}_t + \mathbf{S}^H \mathbf{v}_t$$

$$= \mathbf{RAb}_t + \mathbf{n}_t$$

15   where ^{H} denotes the Hermitian operator and $\mathbf{R}$ is defined by:

$$\mathbf{R} = \begin{bmatrix} 1 & \cdots & \rho_{1,n_T} \\ \cdots & \cdots & \cdots \\ \rho_{1,n_T} & \cdots & 1 \end{bmatrix}$$

20   where

$$\rho_{i,j} = <s_i, s_j>, \quad \rho_{i,j} = \rho^*_{j,i}, \text{ and}$$

with the inner product operation $<s_i, s_j>$ signifying $\sum_{k=1 \text{ to } n_R} (s_{ik} s^*_{jk})$ and * denoting the

complex conjugate.

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From a mathematical standpoint, Equation 1A shows that the space matched filter 68 executes a matrix multiplication operation on r_t (the received vector), which is a linear combination of the elements of r_t, in order to obtain y_t (the filtered vector). For its part, Equation 1B shows that the elements of y_t are noisy and scaled observations of the elements

5   of b_t (the transmitted vector).

It is also noted that because S is intimately related to H, the space matched filter 68 is assumed to have some knowledge of the channel characteristics, i.e., the columns of H. This is a reasonable assumption, as such characteristics could be measured or estimated

10   directly by the receiver 60 based on observations of the channel 46. Alternatively, the transmitter 20 may be provided with an estimator, which is capable of determining H either partially or completely; knowledge of the channel transfer matrix H could then be supplied to the space matched filter 68. Those skilled in the art will recognize that there is no need for the channel transfer matrix H used in the computations correspond exactly to the actual

15   channel transfer matrix. Advantageous performance can be achieved even if the estimates of the channel state information used by the space matched filter 68 deviate from the actual channel state information. For example, a pre-selected value for the channel transfer matrix H could be used, based on previous measurements or estimates, or based on analytical computations or simplifications. If the elements of H, or the norms of the columns of H,

20   are within, say 10% of their true values, then there will still be advantages to space matched filtering using H.

It is further noted that the ρ_{ij} are correlations between the normalized columns of H, i.e., the columns of S. The correlations ρ_{ij} can be non-zero as the columns of H are not orthogonal.

These correlations give rise to co-antenna interference (CAI). For example, the first output

of the space matched filter 68 is given by:

(Equation 2)

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$$y_{1,t} = A_1 b_{1,t} + A_2 \rho_{12} b_{2,t} + \ldots + A_{n_T} \rho_{1,n_T} b_{n_T,t} + n_{1,t}.$$

Since the desired signal is $A_1 b_{1,t}$, the last $n_T - 1$ terms (namely, $\sum_{j=2}^{n_T} A_j \rho_{1j} b_{j,t}$) are due to CAI.

Hence, one may consider the transmitted symbol corrupted by two sources of noise, the

Gaussian noise term ($n_{1,t}$) and the CAI. The more transmitting antennas there are, the more

10   accurately one may approximate this combined interference for each receive antenna as

being Gaussian distributed with a mean and a variance that is function of the ρ_{ij}, i.e., the

correlation between the columns of **H**. In an embodiment, a separate processing unit can be

provided which would comprise suitable circuitry, software and/or control logic to estimate

this mean and variance.

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Continuing with the description of the receiver 60 in Fig. 2, the elements of the vector y_t at

the output of the space matched filter 68 are fed to a detector 70, which comprises suitable

circuitry, software and/or control logic for computing probabilities on the transmitted

symbols. That is to say, given the filtered vector y_t at the output of the space matched filter

20   68, the detector 70 is operative to compute the probability of each element of the transmit

symbol vector b_t being equal to a particular modulation symbol represented by real (or

complex) components. This can be achieved using maximum likelihood or maximum *a*

posteriori techniques based on the filtered vector y_t and assumptions about the noise

statistics through the channel. For instance, the detector 70 can estimate the mean and

variance of the noise for each possible modulation symbol in the transmit symbol vector b_t.

Based on these estimates, the detector 70 computes, for the kth transmit antenna, a set of M

probability values $\{82_k\}$, where M denotes the number of possible modulation symbols.

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Example: computation of probability values

By way of non-limiting example, assume that the information symbols 10 produced

by the digital data source 22 are bits and that these bits are coded using a FEC

10   encoder 24 to produce coded symbols 12 (which are also bits). The coded bits 12

are then passed to the mapper 26, which performs binary phase shift keying (BPSK)

modulation ($M = 2$), mapping coded bit '0' to modulation symbol '-1' and coded bit

'1' to modulation symbol '1'. These modulation symbols are passed to the DT to

CT unit 42, which performs amplitude shift keying. Let the resultant continuous

15   waveforms be transmitted over $n_T = 3$ transmit antennas, and let the receiver have n_R

$= 20$ receive antennas. Let the amplitude matrix A of the channel transfer matrix H

for one instance of transmission be given by:

$$A \quad = \quad \begin{bmatrix} 2 & 0 & 0 \\ 0 & 2.5 & 0 \\ 0 & 0 & 3 \end{bmatrix}$$

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and the correlation matrix R be given by:

$$R \quad = \quad \begin{bmatrix} 1 & 0.2 & 0.1 \\ 0.2 & 1 & 0.15 \\ 0.1 & 0.15 & 1 \end{bmatrix}$$

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Let it also be assumed that it is equally likely for a particular modulation symbol to

be 1 or −1. Under these circumstances, the mean of the CAI for the first transmit

antenna is:

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(Equation 3A)

$$\mu_1 = \sum_{j=2}^{3} \sum_{k=1}^{2} A_j \rho_{1j} Q_{jk} P[Q_{jk}]$$
$$= 2.5*0.2*1*0.5 + 2.5*0.2*(-1)*0.5 + 3*0.1*1*0.5 + 3*0.1*(-1)*0.5$$
$$= 0$$

Although for BPSK transmission the mean is 0, in general, this may not be the case.

10   The variance of the CAI noise for the first antenna is:

(Equation 3B)

$$\sigma_{r1}^2 = \sum_{j=2}^{3} \left(\sum_{k=1}^{2} \left(A_j \rho_{1j} Q_{jk} P[Q_{jk}] \right)^2 - \left(\sum_{k=1}^{2} A_j \rho_{1j} Q_{jk} P[Q_{jk}] \right)^2 \right)$$
$$= \left((2.5*0.2*1*0.5)^2 + (2.5*0.2*-1*0.5)^2 - (2.5*0.2*(1-1)*0.5)^2 \right)$$
$$+ \left((3*0.1*1*0.5)^2 + (3*0.1*-1*0.5)^2 - (3*0.1*(1-1)*0.5)^2 \right)$$
$$= 0.68$$

15   Assuming that the variance of the AWGN on the channel is 0.5, then the variance of

the overall noise is 1.18 (= 0.5 + 0.68). If the output of the space matched filter 68

for the first transmit antenna is assumed to be, say, 0.5, then one implementation of

the detector 70 would compute probability on the modulation symbol as follows:

$$P(b_1|y_1) = \frac{1}{\sqrt{2\pi\sigma_{x1}^2}} \exp\left(-\frac{1}{2}\left(\frac{y_1 - \mu_1 - A_1 b_1}{\sigma_{x1}}\right)^2\right)$$

$$= \begin{cases} \dfrac{1}{\sqrt{2\pi(1.18)}} \exp\left(-\dfrac{1}{2}\left(\dfrac{0.5-0-2}{\sqrt{1.18}}\right)^2\right) \\[2ex] \dfrac{1}{\sqrt{2\pi(1.18)}} \exp\left(-\dfrac{1}{2}\left(\dfrac{0.5-0+2}{\sqrt{1.18}}\right)^2\right) \end{cases}$$

$$= \begin{cases} 0.845, & \text{if } b_1 = 1 \\ 0.155, & \text{if } b_1 = -1 \end{cases}$$

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Returning now to the description of the receiver 60, the n_T sets of M probability values each

(as produced by the detector 70) are then passed to a channel de-interleaver 72, a

multiplexer 74 and then a de-mapper of demodulator 76. The channel de-interleaver 72

performs basically the inverse operations to those of the channel interleaver 36. However,

the channel de-interleaver 72 acts on n_T streams of M probability values per stream rather

than simply n_T symbol sub-streams. Thus, the channel de-interleaver 72 comprises suitable

circuitry, software and/or control logic for time-rearranging each received stream of sets of

probability values $\{82_1\}$, $\{82_2\}$,..., $\{82_{n_T}\}$ into a de-interleaved stream of sets of

probability values $\{84_1\}$, $\{84_2\}$,..., $\{84_{n_T}\}$.

For its part, the multiplexer 74 acts to reassemble the data into the order which existed at the

output of the mapper 26. To this end, the multiplexer 74 comprises circuitry, software

and/or control logic for combining the de-interleaved sets of probability values $\{84_1\}$,

$\{84_2\}$, $\{84_{n_T}\}$ into a single stream of sets of probability values $\{86\}$. Each set of

probability values in this stream is thus associated with a respective transmitted symbol,

with the probability value occupying a particular position in the set representing the

likelihood of that transmitted symbol acquiring a corresponding predetermined value.

The de-mapper 76 serves to convert each set of M probability values received from the

5   multiplexer 74 into a hard or soft decision about the transmitted coded symbol. In the case

of a hard decision demodulation, the de-mapper 76 may comprise suitable circuitry,

software and/or control logic for selecting, as the best transmitted coded symbol, the

predetermined symbol corresponding to the position in the set {86} which is occupied by

the greatest probability value. Of course, other ways of forming hard decisions will be

10   known to those of ordinary skill in the art. In the case of a soft decision, the de-mapper 76

may comprise suitable circuitry, software and/or control logic for creating a probability on

each particular transmitted coded symbol by performing a weighted sum on the probabilities

from the detector 70. Each element of the sum is weighted by a 1 or 0 depending if the

particular coded symbol is contained in the appropriate bit position of the modulation

15   symbol.

For example, for a soft decision 4-QAM de-mapper, at the ith received symbol instant, the

probability of the $2i$-th transmitted coded symbol being 0 is 1*P[modulation symbol for 00]

+ 1*P[modulation symbol for 01] + 0*P[modulation symbol for 10] + 0*P[modulation

20   symbol for 11]. Similarly, the probability of the $(2i+1)$-th transmitted coded symbol being 0

is 1*P[modulation symbol for 00] + 0*P[modulation symbol for 01] + 1*P[modulation

symbol for 10] + 0*P[modulation symbol for 11]

The output of the de-mapper 76 is fed via link 90 to an FEC decoder 80, which performs the

25   inverse function of the FEC encoder 24 (if such FEC encoder 24 is used). It is noted that if

the FEC decoder 80 is a soft-decision FEC decoder 80, then it is possible to configure the

FEC decoder 80 to generate reliability values on the transmitted coded symbols (also known

as *soft decisions or estimates of the coded symbols probabilities*). Then, the de-mapper 76

can be changed to merge these (shown by link 88 in Fig. 2) with the probability values

5   arriving from the multiplexer 74. Otherwise the de-mapper 76 simply passes its soft output

to the FEC decoder 80 (without iteration). The FEC decoder 80 then provides its output to a

data sink 78 via a link 92.

In an alternative embodiment, shown in Fig. 3, the receiver 60' is equipped with iterative

10   processing between the FEC decoder 80 and a modified detector 70'. Specifically, after

computing the reliability values on the transmitted coded symbols, the decoder 80 can then

reassemble these into reliability values on the transmitted modulation symbols (also known

as soft decisions or estimates of probabilities of the modulation symbols). The reliability

values (shown by link 94 in Fig. 3) are provided to the detector 70'. Given this additional

15   information, the mean and variance of the CAI noise source may then be re-computed by

the detector 70' (using (Equation 3A) and (Equation 3B) above, with different values for the

$P[Q_{jk}]$), which will tend to more accurately reflect the actual mean and variance of the CAI.

To further improve the performance of the considered system, an interference reducing filter

20   can be applied to the output of the space matched filter 68. One example of a suitable

interference reducing filter is a minimum mean square error (MMSE) filter, although other

conventional examples of interference reducing filters (e.g., interference reducing filters

used in speech systems) are suitable and will be known to those skilled in the art. Fig. 4

illustrates a block diagram of a receiver 60'' equipped with this optional functionality. The

25   detector 70 has the same structure as that described in Fig. 2, however it operates on the

output of an interference reducing filter 71. In one embodiment, the interference reducing

filter 71 is an MMSE filter and performs a function defined by:

$$\mathbf{M}_y = \left(\mathbf{R} + \sigma_{AWGN}^2 \mathbf{A}^{-2} \right)^{-1}.$$

5

The output of the interference reducing filter 71 is thus given by:

$$\begin{aligned} \mathbf{z}_t &= \mathbf{M}_y \mathbf{y} \\ &= \mathbf{M}_y \mathbf{R} \mathbf{A} \mathbf{b}_t + \mathbf{M}_y \mathbf{n}_t \end{aligned}$$

10   which can be written as:

$$\mathbf{z}_t = \mathbf{C} \mathbf{b}_t + \mathbf{w}_t ,$$

where $\mathbf{C} = \mathbf{M}_y \mathbf{R} \mathbf{A}$. One can write the output of the interference reducing filter 71 for the

15   first transmit antenna as:

(Equation 4)

$$z_{1,t} = C_{11} b_{1,t} + C_{12} b_{2,t} + ... + C_{1 n_T} b_{n_T, t} + w_{1,t} ,$$

20   where the desired signal is $C_{11} b_{1,t}$, the last $n_T - 1$ terms are due to (reduced) CAI ($\sum_{j=2}^{n_T} C_{1j} b_{j,t}$)

and $\mathbf{w}_t$ is assumed to be a Gaussian noise source. Clearly, (Equation 4) is very similar in

form to (Equation 2). Hence, it is possible to calculate the mean and variance in a similar

fashion as before. Then the detector 70 can compute the probability values on the

transmitted symbols and pass these on to the channel de-interleaver 72, multiplexer 74, de-mapper 76 and subsequently to the FEC decoder 80 as has already been described.

Those skilled in the art will appreciate that convenience of mathematical expression has led

5   to the design of a system that is optimized for quasi-static, multi-path fading with little or no

intersymbol interference (ISI) between symbols (i.e., the fading is constant over a

transmitted multi-symbol packet). While the performance of the present invention may be

affected by the degree to which the fading, multi-path and noise conditions divert from

those described in the foregoing, it is nonetheless expected that advantage can still be

10   derived from the invention under a wide range of conditions. In other words, the conditions

and assumptions described herein for purposes of simplifying the mathematical exercise are

not to be construed as limitations upon the usefulness, applicability or scope of the present

invention.

15   For example, the techniques described herein find utilization in the presence of ISI induced

by a frequency-selective fading channel. One example of such utilization (e.g., in a

TDMA/FDMA uplink/downlink system) involves the use of an ISI reducing equalization

filter at the front end of the receiver, in which case the residual ISI contributions will be

ignored by the detector 70 and treated as an additional source of noise. Another example of

20   such utilization (e.g., in case of the CDMA uplink transmission) involves the use of a

RAKE receiver to collect all the energy of the power delay profile of the signal at the front

end of the receiver, in which case one would again proceed as previously described,

utilizing the detector 70 and (possibly iterative) post processing after the RAKE receiver

pre-processing.

25

The present invention is also applicable in case of multiple users sharing the same

communication channel, e.g., for uplink transmission from the mobile to the base station.

Specifically, the receiver described in Figs. 2, 3 and 4 can be applied directly to the case of

TDMA, FDMA and orthogonal CDMA based multi-user communication systems, as the.

5   users can be separated due to their presence in different time/frequency/code dimension

(i.e., the users are orthogonal). However, for non-orthogonal CDMA systems that employ

multiple antenna technology, the following considerations apply.

Specifically, Fig. 5 illustrates a schematic block diagram of a transmission scenario for K

10   users on a CDMA network. Each user has a conventional single user transmitter 20^1, ...,

20^K with single or multiple antennas, depending on the type of mobile unit (e.g., laptop,

cellular phone, PDA, etc.) Thus, each of the transmitters 20^1, ..., 20^K comprises a

respective one of K digital data sources 22^1, ..., 22^K, resulting in the production of K

streams of information symbols that pass through respective optional FEC encoders 24^1, ...,

15   24^K. The FEC encoders 24^1, ..., 24^K, if used, include suitable circuitry, software and/or

control logic for adding redundancy and/or memory to each stream of information symbols

by encoding them into a respective sequence of coded symbols.

The K streams of coded symbols are fed to a respective plurality of mappers 26^1, ..., 26^K,

20   each of which includes suitable circuitry, software and/or control logic for translating the

coded symbols in the respective stream of coded symbols into a respective stream of

modulation symbols. (In the specific example being considered here, each mapper includes

a respective user's spreading CDMA sequence.) Each stream of modulation symbols is fed

to a respective one a plurality of de-multiplexers 32^1, ..., 32^K, each of which includes

25   suitable circuitry, software and/or control logic for separating the respective stream of

modulation symbols into a respective plurality of modulation symbol sub-streams. Each de-

multiplexer 32^1, ..., 32^K is connected to a respective channel interleaver 36^1, ..., 36^K, which

includes suitable circuitry, software and/or control logic for changing the order in which

modulation symbols appear within each of the respective modulation symbol sub-streams.

5   The output of each channel interleaver 36^1, ..., 36^K is therefore a respective plurality of

modulation symbol sub-streams.

The set of modulation symbol sub-streams produced by each of the channel interleavers 36^1,

..., 36^K are fed to a respective plurality of transmit antennas via a respective DT-to-CT unit

10   (not shown). The DT-to-CT unit modulates each modulation signal sub-stream in

accordance with a CDMA chip sequence for that symbol. Specifically, the continuous-time

waveforms for the modulation symbol sub-streams produced by channel interleaver 36^1 are

fed to a transmit antennas $40^1{}_1$, $40^1{}_2$,..., $40^1{}_{n^1_T}$, while the continuous-time waveforms for

the modulation symbol sub-streams produced by channel interleaver 36^1 are fed via a

15   respective transmit antennas $40^2{}_1$, $40^2{}_2$,..., $40^2{}_{n^2_T}$, and so on. Thus, it is seen that the kth

user is associated with n^k_T transmit antennas.

Fig. 6 illustrates the schematic block diagram of a receiver 60* (e.g., at the base station)

having a multi-user detector / decoder. It is assumed that the scenario of Fig. 5 applies., i.e.,

20   there are K distinct users transmitting their data to the base station over the same

communication channel, while utilizing n^k_T transmit antennas respectively. The receiver

60* includes n_R receive antennas. Advantageously, n_R will be greater than one. The output

of each of the n_R receive antennas is a respective continuous-time waveform, which is

provided to a respective portion of a CT-to-DT unit 66*. The CT-to-DT unit 66 comprises

25   suitable circuitry, software and/or control logic for discretizing each continuous-time

waveform to produce a sequence of numerical values therefrom. In particular, the CT-to-

DT unit 66* is operative to demodulate the continuous-time waveforms $64_1, 64_2,..., 64_{n_R}$

using a demodulation scheme that is based on CDMA chips, i.e., it operates at the chip rate

of the DT-to-CT unit in the transmitter (not shown in Fig. 5). The received signal (after the

5   CT-to-DT unit 66*) can be represented as follows, assuming that the conversion has been

done on the CDMA chip level:

$$
\begin{bmatrix} \mathbf{H_1} & 0 & \cdots & 0 \\ 0 & \mathbf{H_2} & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & \mathbf{H_K} \end{bmatrix}
\begin{bmatrix} \left[\mathbf{w}_1^1, \mathbf{w}_1^2, ..., \mathbf{w}_1^{n_T^1} \right] & 0 & \cdots & 0 \\ 0 & \left[\mathbf{w}_2^1, \mathbf{w}_2^2, ..., \mathbf{w}_2^{n_T^2} \right] & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & \left[\mathbf{w}_K^1, \mathbf{w}_K^2, ..., \mathbf{w}_K^{n_T^K} \right] \end{bmatrix}
\begin{bmatrix} \mathbf{b_1} \\ \vdots \\ \mathbf{b_K} \end{bmatrix}
$$

where

- sub-matrix $\mathbf{H}_k$ is an n_R x n^k_T matrix reflecting the channel attenuation factors

10   between the n^k_T antennas of the kth user and the n_R receive antennas of the base

station,

- the spreading sequence of length N, utilized by the kth user at the kth user's j-th

transmit antenna is represented by a column vector $\mathbf{w}^j_k$ of size N x 1,

- the data symbols transmitted by the kth user are given by a vector $\mathbf{b}_k$ of size n^k_T x 1.

15

The CT-to-DT unit 66* is followed by a space matched filter 68*. The space matched filter

68* can be implemented as a bank of CDMA matched filters or a suitable CDMA multi-

user detector. The basic idea of the space matched filter 68* is, for each kth user, to

correlate (or match) the received vector with a "vector of chip amplitudes" corresponding to

20   each of that user's transmit antennas $40^k_1, 40^k_2,..., 40^k_{n^k_T}$. The "vector of chip amplitudes"

corresponding to a given transmit antenna of a given user is indicative of a channel

characteristic associated with that transmit antenna. In one embodiment, the "vector of chip

amplitudes" corresponding to a given transmit antenna of a given user is indicative of the

degree to which the signal from that transmit antenna of that user has been spread out

amongst the n_R receive antennas. In the context of equalization of a multi-input multi-

5   output channel, the "vector of chip amplitudes" corresponding to the jth transmit antenna of

the kth user is ideally the jth normalized column of the kth channel transfer sub-matrix $\mathbf{H}_k$.

Mathematically, an equation similar to Equation (1B) applies to each kth user . Specifically,

denoting $\mathbf{y}_k$ as the filtered vector for the kth user (i.e., at the output of the space matched

10   filter 68*) and ignoring the subscript "t", one has:

(Equation 5)

$$\mathbf{y}_k = \mathbf{R}_k \mathbf{A}_k \mathbf{b}_k + \mathbf{n}_k$$

15   for the amplitude matrix $\mathbf{A}_k$ of sub-matrix $\mathbf{H}_k$, where ^{H} denotes the Hermitian operator and

$\mathbf{R}_k$ is defined by:

$$\mathbf{R}_k = \begin{bmatrix} 1 & \cdots & \rho^k_{1,n^k_T} \\ \cdots & \cdots & \cdots \\ \rho^k_{1,n^k_T} & \cdots & 1 \end{bmatrix}$$

20

where

$$\rho^k_{i,j} = <\mathbf{s}^k_i, \mathbf{s}^k_j>, \quad \rho^k_{i,j} = \rho^{k*}_{j,i}$$

with $\mathbf{s}^k_i$ being the ith column of sub-matrix $\mathbf{H}_k$ and with the inner product operation $<\mathbf{s}^k_i, \mathbf{s}^k_j>$

25   signifying $\sum_{m=1 \text{ to } n_R} (\mathbf{s}^k_{i,m} \mathbf{s}^{*k}_{j,m})$ and * denoting the complex conjugate.

Continuing with the description of the receiver 60*, the output of the space matched filter

68* leads to a soft decision detector 70* and, optionally, an interference reducing filter 71*.

After the detector 70* (and the optional interference reducing filter 71*), the data are treated

in parallel (i.e., separately) by a bank of de-interleavers 72^1, ..., 72^K, followed by a bank of

5   multiplexers 74^1, ..., 74^K and, if used, a bank of FEC decoders 80^1, ..., 80^K for users 1, 2,

..., K. Hence, the mathematical treatment applied in the single user case be applied as

before to estimate the symbols transmitted from each transmit antenna of every user. The

output of each decoder 80^1, 80^2, ..., 80^K is fed to a respective data sink 78^1, 78^2, ..., 78^K,

one for each of the K users.

10

It is noted that if the FEC decoders 80^1, ..., 80^K are soft decision decoders, then it is

possible to configure these decoders to generate reliability values (shown by links 88^1, ...,

88^K in Fig. 6) on the transmitted coded symbols. In this case, each of the de-mappers 76^1,

..., 76^K can be designed to merge these reliability values with the probability values arriving

15   from the respective multiplexer 74^1, ..., 74^K as .

In an alternative embodiment, the receiver 60* is equipped with iterative processing. After

computing the reliability values on the transmitted coded symbols, the FEC decoders 80^1,

..., 80^K can then reassemble these reliability values into reliability values on the transmitted

20   modulation symbols. These reliability values (also known as soft decisions or probability

estimates of the transmitted modulation symbols) are provided to the detector 70* via

respective links 94^1, ..., 94^K. Given this additional information, the mean and variance of

the CAI noise source may then be re-computed by the detector 70* (using equations

analogous to (Equation 3A) and (Equation 3B), above, with different values for the $P[Q_{jk}]$),

25   which will tend to more accurately reflect the actual mean and variance of the CAI.

Those skilled in the art will of course appreciate that among the many variations that

achieve the same goal of overall data recovery, one can merge some or all of the CT-to-DT

unit 66*, the space matched filter 68* and the detector 70*, or one can change the order in

5   which the space matched filter 68* and the detector 70* appear in the receiver 60*.

Those skilled in the art will also appreciate that the present invention need not be restricted

to use in wireless systems. In particular, it is contemplated that the present invention can

also be useful to solve detection problems in those situations where the channel over which

10   data is transmitted can be modeled by way of a channel transfer matrix of the type described

herein above. Specifically, other examples of such areas of application include, but are not

limited to (i) wireline data transmission when data are affected by channel cross-talk; (ii)

data readout from CD/DVD/other optical recording medium when multiple lasers and

detectors are used concurrently; (iii) data read-out from a magnetic disk or magnetic tape

15   when multiple read-heads are used; and (iv) deep sea data communications. In each of the

above scenarios, there exists channel cross-talk due to multiple concurrent

transmissions/receptions within the same bandwidth and physical area. The channel

transfer matrix can describe the effect of transmission over such a channel and the effects

can be exploited to advantage through the use of a space matched filter.

20

While specific embodiments of the present invention have been described and illustrated, it

will be apparent to those skilled in the art that numerous modifications and variations can be

made without departing from the scope of the invention as defined in the appended claims.